

# **Packaging of Microwave Integrated Circuits Operating Beyond 100 GHz**

Erik Daniel, Vladimir Sokolov, Scott Sommerfeldt, Jim Bublitz, Kim Olson, Barry Gilbert

Mayo Foundation,  
Rochester, MN 55905, USA

Lorene Samoska  
Jet Propulsion Lab  
Pasadena, CA 91109

David Chow  
HRL Laboratories  
Malibu, CA 90265

## **Abstract**

Several methods of packaging high speed (75-330 GHz) InP HEMT MMIC devices are discussed. Coplanar wirebonding is presented with measured insertion loss of less than 0.5 dB and return loss better than -17 dB from DC to 110 GHz. A motherboard / daughterboard packaging scheme is presented which supports minimum loss chains of MMICs using this coplanar wirebonding method. Split waveguide block packaging approaches are presented in G-band (140-220 GHz) with two types of MMIC-waveguide transitions: E-plane probe and antipodal finline.

## **I. Introduction**

In recent years, improvements in integrated circuit technologies have made robust, reproducible active monolithic microwave integrated circuits (MMICs) operating above 100 GHz a reality. Such circuits enable applications in high bandwidth communications, passive remote sensing, gas analysis, and millimeter wave astronomy. Although techniques for characterization and packaging of these mm-wave MMICs can often be scaled from lower frequency approaches, for frequencies above 100 GHz their implementation can become the primary factors limiting performance and subsequent system insertion. A significant amount of work has been performed in the area of packaging and testing passive electrical devices operating up to extremely high frequencies (in the few THz) [1-2]. However, active devices provide additional challenges. This paper presents some examples of packaging of MMICs fabricated in an 0.1  $\mu\text{m}$  InP HEMT process at HRL Laboratories with operating frequencies in the 75-330 GHz range. First, we discuss wirebond performance in this frequency range, as this is relevant to all packaging schemes presented here (as well as many others). Then, two packaging approaches are described. The first, a motherboard / daughterboard approach, is aimed at chaining many die with minimum loss while maintaining good DC delivery. The specific implementation demonstrated is for rapid prototyping in a laboratory setting. The second packaging approach is more conventional, interfacing the MMICs with standard mm-wave waveguide in split-block fixtures.

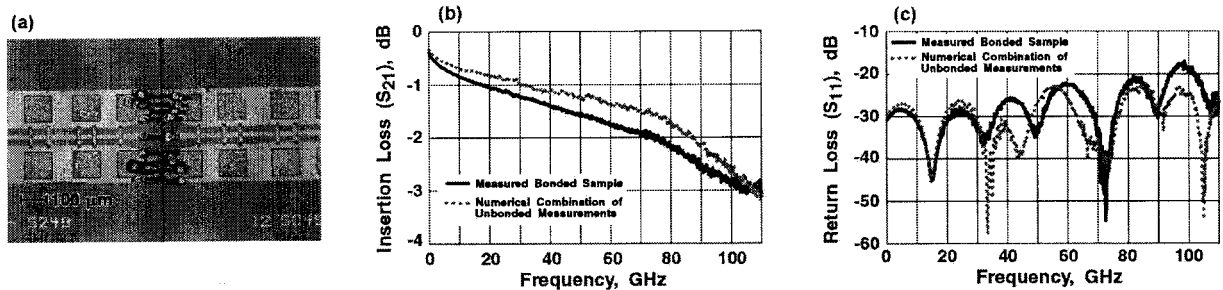
## II. Wirebond Performance

The packaging schemes described below rely on wirebonding of the MMIC die to other components. Although this is common practice in the mm-wave circuit field, and published literature describing wirebond interfaces operating up to 40 GHz is available [3-4], we are aware of little published work characterizing such bonds up to or above 100 GHz. Here we present results of a study of coplanar wirebond interfaces between two 50  $\Omega$  coplanar waveguide with ground (CPWG) transmission lines fabricated in the same InP HEMT process as the MMICs described in this paper.

Each CPWG line is 1.83 mm long and contains closely spaced vias to the metallized back side of the 2 mil thick InP substrate as well as air bridge connections between the two coplanar grounds to suppress higher order modes. Measurements were performed using an HP8510XF vector network analyzer (VNA) with GGB Industries microwave probes allowing single-sweep measurement from 45 MHz to 110 GHz. An LRRM calibration was performed to the probe tips using a GGB CS-5 calibration substrate and Cascade Microtech's WinCal<sup>TM</sup> software. Each line was individually measured before bonding was performed, in addition to the measurements performed after bonding.

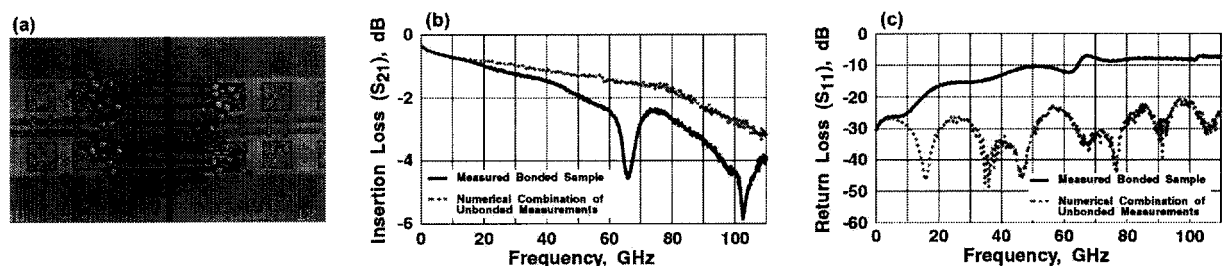
The bonding procedure used for preparing these samples (and for high frequency MMIC bonding in general) is quite different from standard production wedge or ball bonding. The goal is to provide minimum length gold bonds very close to the substrate surface in such geometry as to minimize inductive or capacitive discontinuities at the interface. In order to minimize the bond length, it is sometimes necessary to remove the excess bare InP substrate left between the edge of the pads and the die edge after scribing. While additional dicing or scribing steps can be performed, it is exceedingly difficult to do this without chipping or otherwise damaging the extremely fragile 2 mil thin InP die. Our typical procedure is to grind down the die edge on a polishing turntable. After this step, the die are epoxied down onto a gold-plated handling substrate, with the polished die edges butted together. Then, 0.5 mil diameter gold wires are manually cut to length and bonded in place using a manual thermocompression bonder. Note that care is taken to ensure minimal spread of the bond feet on the center conductors into the signal-ground gap to minimize capacitive discontinuity. Also, multiple bond wires are provided on each ground with one very close to the edge near the signal trace in order to minimize inductive discontinuity.

Photographs and measured data for a sample prepared in this manner are shown in Figure 1. Figures 1(b)-(c) show measured S-parameters for the bonded sample as well as the numerical chain combination of the measured S-parameters of the individual lines – an approximation of a “perfect” bond. The incremental insertion loss contributed by the bond is less than 0.5 dB across the entire measurement range, and the return loss is less than -20 dB across nearly the entire range, edging up to -17 dB at 100 GHz. In short, excellent electrical performance is provided up to at least 110 GHz and likely much higher.



**Figure 1. (a) Photograph and (b), (c) measured S-parameters of bonded CPWG transmission lines with minimum length bonds.**

As the processing required to remove the excess bare InP between the pad edges and the die edge can be considerable, it is natural to ask to what extent this improves the electrical performance. Photographs and measured data taken on a sample prepared without the removal of the excess InP (roughly 300 microns total between the pads) are shown in Figure 2. The measured S-parameters (Figures 2(b)-(c)) show significantly degraded insertion and return loss above 40 GHz. Observation of the time domain step response representation of  $S_{11}$  (analogous to a time domain reflectometry measurement, not shown here) indicates somewhat more distortion and a significant inductive reflection at the bond interface. It is somewhat surprising that the degradation is so severe, given that the set of bond wires replicates the top-side coplanar environment fairly well. We speculate that the tendency of the bond wires to pull up slightly from the surface and the lack of vias or air bridge ties in the bond region are responsible for the parasitic inductance and resonances evident.



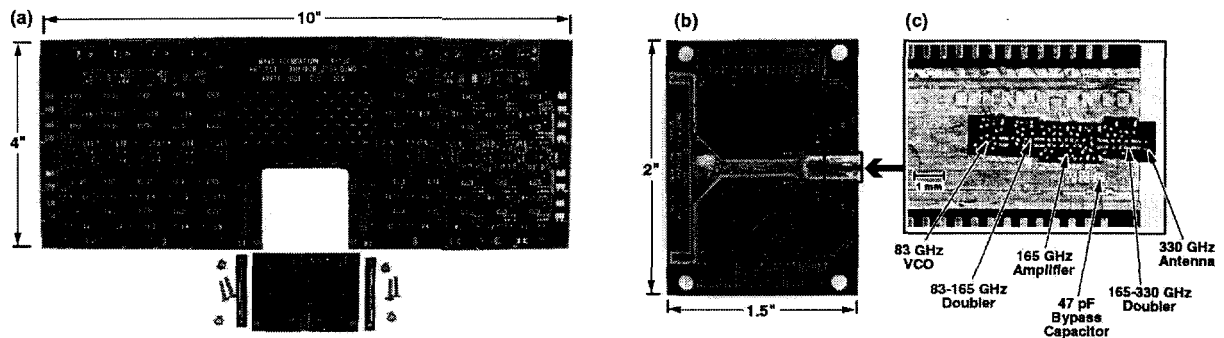
**Figure 2. (a) Photograph and (b), (c) measured S-parameters of bonded CPWG transmission lines without excess substrate removed between pads.**

### III. Motherboard / Daughterboard Packaging

The most commonly used methods of packaging small circuits with inputs and outputs operating at more than 100 GHz utilize waveguide interfaces. As will be discussed in the next section, the key issue with waveguide packaging schemes is the design of a transition from the waveguide modes to the CPWG or microstrip environments typically used on the integrated circuit chips. Here we present an alternative packaging scheme that can be used for systems in which many MMIC die are to be chained end-to-end, in which these

waveguide transitions are avoided, thereby minimizing the loss and volume associated with the packaging. This situation is not uncommon, as low yield of extremely high frequency, high performance devices is common, making monolithic integration of such a chain of MMICs difficult or impossible. In addition, in a development phase, it may be desirable to evaluate each component individually before chaining all devices together, not only due to yield issues, but also as accurate simulation and modeling of such high frequency circuits is extremely challenging.

The specific motherboard / daughterboard packaging approach described here (shown in Figure 3(a)) was designed for development and test of high frequency sources comprised of a chain of InP HEMT MMICs. For example, a packaged 330 GHz source is shown in Figures 3(b)-(c), consisting of an 82.5 GHz voltage controlled oscillator (VCO) followed by an active doubler, a 165 GHz amplifier, a second active doubler, and finally, a 330 GHz quasi-Yagi antenna[5-9]. Several design considerations led to the implementation shown. Many different candidate chains of MMIC die were to be considered, so the design could not be optimized for a single MMIC chain – versatility was required. Independent control of each DC bias was desired for optimization. Minimization of the signal loss along the chain was deemed critical. It was assumed that there would be no external input (with the VCO providing the initial input to the subsequent stage) and that the output would either be to a high frequency microwave probe for laboratory testing, or antenna coupling to free space (as shown in Figure 3(c)).



**Figure 3. Photographs of (a) Motherboard with elastomer interconnects and daughterboard, (b) close-up of daughterboard with 330 GHz source MMICs, and (c) close-up of MMIC die in daughterboard cavity.**

The daughterboard is simply a carrier for the MMIC die and some associated passives (e.g. bypass capacitors). It consists of a thick (1/8") copper plate covered with a layer of dielectric prepreg (2.5 mil polyimide) and core laminate (4 mil thick polyimide). A channel is milled through the dielectric layers, slightly into the copper base, the side walls of which are plated to provide a continuous wrap-around ground from the copper base to a ring around the cavity on the top surface of the dielectric. A 5 mil thick gold-plated copper shim is epoxied into the channel, on which the MMIC die as well as wirebondable bypass capacitors are placed. The shim provides a very flat surface on which the die can sit, and it also juts out slightly from

the end of the copper base, providing additional support to the antenna die. A single gold-plated copper layer on top of the dielectric layers provides fan-out of up to 40 traces designed for delivering DC or low-speed (1 GHz) signals to the die from pads along the two shorter edges of the daughterboard. It was not anticipated that this many DC or low speed signal lines would be needed for any single MMIC string, but a large density was required in order that a wide range of chip designs and arrangements could be accommodated while still maintaining the flexibility of providing independent control of each DC connection with good power bypass capability (which is critical for stability).

The daughter board is mounted face-up onto the underside of the motherboard using quick-disconnect vertical elastomer interconnects. The motherboard contains 40 independent simple linear DC supplies which can be used to supply the daughterboard traces with DC bias. Alternatively, these supplies can be disabled and bias can be brought in externally, such that the motherboard merely provides a convenient means for connecting them to the daughterboard. Additional bypass capacitance can also be placed on the motherboard as needed. We note that the MMIC die and associated high frequency interconnects consume a tiny fraction of the motherboard / daughterboard volume, with the rest dedicated to power or low speed signal delivery due to our desire for flexibility. Certainly, one could optimize a similar design specific to a given chipset to consume much less volume. Each high frequency MMIC-MMIC interface in the assembly shown in Figure 3 was prepared using the method described in the above section for bonding of two transmission lines together. After this bonding was completed, the bypass capacitors were added and DC bond connections made using a commercial gold ball bonder.

Although all die were DC functional immediately following packaging, the 330 GHz source depicted in Figure 3 was found to only emit a few  $\mu\text{W}$  of power (compared with the 0.1-1 mW expected), and it was later determined one or more of the MMIC die had failed during or just before high speed testing. For multiple, reasons, we have not attempted to prepare another such 330 GHz source. Nonetheless, this packaging scheme has been used with success: as described elsewhere [8], detailed tests of the 165-330 GHz doubler MMIC used in the 330 GHz source chain were performed using die mounted in a similar package. Certainly, one disadvantage of this packaging method is that it is quite difficult to replace one die in the chain if it is malfunctioning, and it is practically impossible to reconfigure the die once they are committed to a given package. We have therefore moved towards waveguide block packaging methods (described in the following section) for continuation of this work as it allows rapid reconfiguration and replacement of chain members. However, we maintain that a similar packaging scheme still has significant advantages for compact low-loss packaging of MMIC chains with minimum interconnection loss and reflection when the constituent die are fairly well-characterized and robust. In addition, we point out that the shim mounted in the daughterboard can be removed with die and bypass capacitors intact such that transfer to another (e.g. waveguide) package is possible after testing on the daughterboard.

## IV. Waveguide Block Packaging

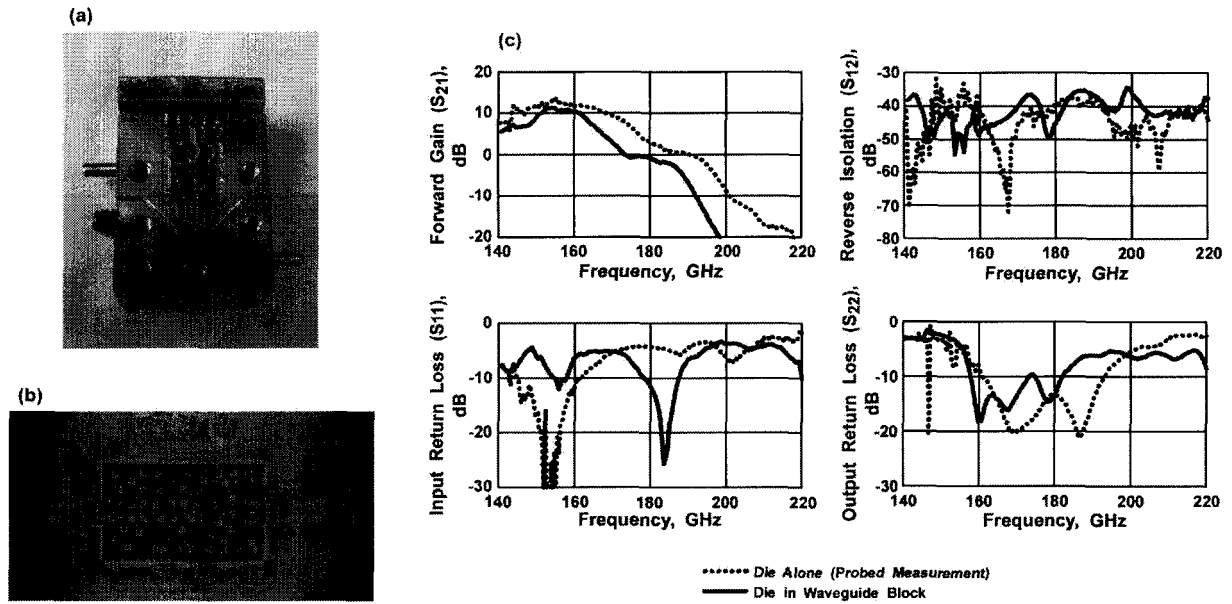
In the general case of a two-port MMIC (or two-port cascade of MMICs), system insertion requires a robust package which is compatible with standard mm-wave interfaces. At frequencies above 100 GHz rectangular metal waveguide is most often employed. Consequently, the package must include a transition between waveguide and the on-chip transmission line of the active device (usually CPWG or microstrip). To ensure minimum loss through the transitions, careful attention must be paid to their design. Several possibilities for the transition circuit present themselves based on scaling of lower frequency, microwave designs. Presented below are the results for two such approaches scaled to G-band (140 to 220 GHz). The first approach utilizes E-plane probe transitions fabricated on alumina substrates, while the second approach utilizes tapered antipodal finline etched on Cuflon<sup>®</sup> (copper on Teflon<sup>®</sup> from Polyflon). Both types of transition circuits are mounted within waveguide channels (albeit with different orientation) milled into solid metal split-block fixtures.

### *E-plane Probe Transitions*

Planar probes have been extensively used at mm-wave frequencies for transitioning from rectangular waveguide to microstrip [10-12]. These planar probe circuits can be printed on any good microwave dielectric such as alumina. The probe itself is basically a metallized patch antenna that penetrates the waveguide at right angles through an opening in the broad wall and sits entirely within the waveguide. One end of the patch is connected to a microstrip line through an impedance matching section to connect to the MMIC outside the waveguide. As such the connecting microstrip line is perpendicular to the longitudinal direction of the waveguide. For minimum loss the connecting microstrip must be as short as possible. This dictates that the waveguides at input and output must provide 90 degree (E-plane) bends. The dimensions of the patch, its penetration into the guide, and its distance from a backshort in the waveguide (for impedance matching to the waveguide) are all part of the design of the transition. The plane of the patch and its supporting substrate can be either perpendicular to the narrow wall [11], or parallel to it [12]. In the latter case it is called an E-plane probe transition. E-plane probe transitions printed on alumina and Teflon<sup>®</sup> substrates exhibit insertion loss of less than 0.9 dB and 0.7 dB, respectively, at 100 GHz based on insertion loss measurements of back-to back transitions [12].

We show an example of the use of E-plane probe transitions fabricated on 2-mil alumina for characterizing an InP 165 GHz medium power MMIC amplifier (one of the HRL die mentioned earlier) in a split-block waveguide fixture. The split block, which is machined out of two blocks of metal (gold-plated brass, in this case), is a common approach for packaging mm-wave devices and circuits. Waveguide channels, as well as cavities for active device mounting, and mounting of printed circuit boards (PCB) for bias distribution, can be milled into each half of the block. Figure 4(a) shows the bottom half (base) of such a block with mounted MMIC, E-plane probe transitions, and PCB containing the bias network. All

components are mounted with silver-laden epoxy. Also seen in the photo is the bottom half of the G-band waveguide channel. The block is machined so as to split the waveguide down the middle of the broad wall. Half the channel resides in the base with the other half in the mating cover. Both input and output waveguide channels complete 90 degree turns prior to interfacing with the E-plane probe transitions (seen as two 45 degree bends in Figure 4(a)). Note the backshort (end of milled channel) that is permanently machined into the block in Figure 4(b). Bias for the MMIC is fed from the printed circuit board and provides a common drain voltage and provision for separately adjusting all gate voltages. For stabilizing the high gain transistors at low frequencies, bypass chip capacitors ( $\sim 0.1 \mu\text{F}$ ) and series ferrite beads are used on the PCB. Smaller, 47 pF chip capacitors are mounted directly adjacent to the MMIC.



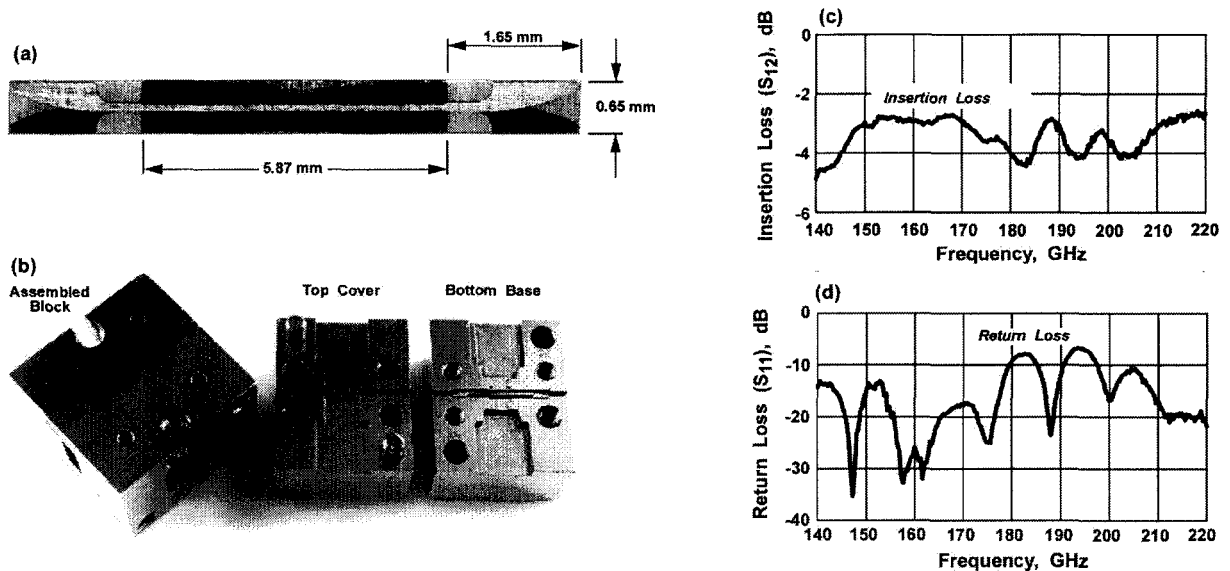
**Figure 4.** Photographs of (a) full waveguide block (bottom half) and (b) close-up of G-band amplifier with E-plane probe transitions. (c) Measured S-parameter data of amplifier mounted in waveguide block along with measured data on bare MMIC die (using microwave probes).

The S-parameters of the packaged amplifier were measured across the 140-to-220 GHz frequency range (solid lines) using an HP 8510C VNA with extender heads (downconverters) from Oleson Microwave Labs for extending the VNA capability into G-band. Calibration was accomplished using a G-band waveguide calibration kit. Figure 4(c) shows the log magnitude plots for all four S-parameters of the packaged amplifier. A peak gain of 11 dB is obtained at around 155 GHz, and “useful” gain of more than 6 dB is obtained from around 140 to 170 GHz. For comparison, also shown is the corresponding data taken with GGB high frequency probes on bare die with the reference plane established at the probe tips. The forward gain of the packaged amplifier is seen to be degraded by as much as 6 dB in the 160 to 180 GHz band. Similarly, the return loss is generally poorer for the packaged amplifier. The extra loss is due to the unavoidable dissipative loss of the transitions, wire bond

interfaces, and limitations of the E-plane probe implementation at G-band (precision placement, backshort optimization, etc.). Although direct measurement of the E-plane probe loss at G-band (e.g., through a measurement of a back-to-back structure) is not available, based on the insertion loss results of Figure 4(c) and published data at lower frequencies [12], we estimate the loss per transition to be at least 1.1 dB at 165 GHz.

### *Finline Transitions*

As an alternative to the E-plane transitions on alumina, another viable approach for interfacing mm-wave MMICs with waveguide is to use a planar transition based on tapered antipodal finline. This type of transition was first described by van Heuven [13] who demonstrated it in the upper microwave range (17.5 to 26 GHz in two waveguide bands) using fused quartz for the substrate. Although quartz is an excellent microwave dielectric material, it is quite fragile, and a more robust solution is to use a soft substrate such as Teflon®[14]. The transition essentially consists of three cascaded transmission lines: a tapered antipodal finline section; a balanced “stripline” section; and finally the unbalanced microstrip (see Figure 5(a)). The plane of the fin-supporting substrate, and the metallized fins, are centered between the narrow walls of the waveguide. The grounded sides of the fins terminate at, and are short-circuited by each broad wall. The dielectric material is kept as thin as possible (limited by mechanical constraints) to minimize unnecessary dielectric loading of the waveguide and to ensure single mode excitation of the microstrip. We have developed a scaled version of such a transition for application in G-band. Its performance in terms of a back-to-back test structure is presented next. To our knowledge, this is the first demonstration of van Heuven’s transition above 110 GHz.



**Figure 5.** Photographs of (a) back-to-back G-band finline transitions, and (b) same structure mounted in waveguide block, along with measured (c) insertion loss and (d) return loss.

Figure 5(a) shows an example of a copper-metallized test structure for G-band operation etched on 2-mil-thick Cuflon<sup>®</sup>. The structure consists of two back-to-back transitions with an interconnecting microstrip line. In the figure the dark portion of the pattern is actually the backside metallization which, in the center of the structure, serves as the ground plane for the microstrip line. In the tapered section the taper opens up to the full height of the G-band waveguide (0.65 mm). One practical advantage of this transition is that it is aligned longitudinally with respect to the waveguide, and is therefore consistent with an in-line physical configuration for any two-port application.

For mounting and testing of the transitions a split-block approach is used as before. Figure 5(b) shows the assembled split-block with the G-band waveguide aperture, the top cover, and the base with the test structure mounted in the milled channel. Provision is made in the channel to support the Cuflon<sup>®</sup> card by its edges, so that when the two halves of the block are assembled the edges of the card are clamped between the cover and base. Cavities are also provided in both halves of the block to accommodate PCBs for bias distribution in future active device assemblies. The design of the transitions was basically scaled from previous lower frequency circuits [14] and optimized for operation in the lower half of the waveguide band. The taper profile is an empirically-derived profile and uses arcs of a circle for the taper shape. Electrically, the length of the taper is about 1.5 wavelengths for a center frequency of 165 GHz.

Measurements at G-band were performed using the HP 8510C VNA and the Oleson extender heads mentioned previously. Figures 5(c) and 5(d) show measured results for the insertion loss and return loss, respectively, for the test circuit described above. From about 150 GHz to 170 GHz the insertion loss is less than 3 dB, while the return loss is better than -12 dB across this band. From an independent calculation (Agilent's ADS; multilayer TLines) the loss of the 5.9 mm-long microstrip line alone is about 0.5 dB. Therefore, the loss per transition is estimated to be no worse than 1.3 dB. Note that at 220 GHz the return loss is -20 dB and the insertion loss is likewise less than 3dB. Consequently, there is potential for full band operation with insertion loss per transition of less than 1.3 dB. Work is currently underway to improve the performance of the transition to cover a wider band, and to insert a G-band MMIC amplifier in place of the microstrip line.

Finally, the relative merits of each transition approach are summarized in Table I with some additional clarification presented in the following. Both transitions have potential for covering full waveguide bands with good return loss performance (subject to the usual gain-bandwidth tradeoff). For the E-plane probe the backshort and its precise location with respect to the probe is critical for optimized electrical performance. The width of the 50  $\Omega$  line on alumina is more compatible for interfacing with typical MMICs than the wider 50  $\Omega$  line on Cuflon<sup>®</sup>. Generally, the finline circuit is more tolerant of small misalignments within the waveguide and is directly compatible with an in-line port-to-port configuration. Although both types of transitions require double-sided processing, the front-to-back alignment is more critical in the finline transition. We note that both types of transitions could be fabricated either in a thin film process (e.g. gold on alumina) or a laminate process

(e.g., copper on Teflon<sup>®</sup>), but there are limitations. Thin film processes usually require a hard, brittle substrate which is mechanically problematic for the finlines as the fins must be clamped by the waveguide block. On the other hand, laminate processes typically cannot allow such fine features or dimensional tolerances as thin film processes. In addition, for long term reliability and ease of wirebonding, gold plating of the copper traces (which is not trivial) is required.

**Table I. Comparison of Waveguide Transition Approaches for >100 GHz Operation**

<b>Characteristics</b>	<b>E-Plane Probe</b>	<b>Antipodal Finline</b>
Insertion loss per transition	1.1 dB @ 165 GHz on alumina	<1.3 dB @ 165 GHz on Teflon <sup>®</sup>
Bandwidth	Full waveguide band possible	Full waveguide band possible
Orientation w.r.t. waveguide	90°	In-line
Unique circuit requirements	Backshort optimization; 90° bend in waveguide	Fin grounding at waveguide broad walls;
Width of 50 $\Omega$ microstrip	1.9-mils on 2-mil alumina	5.9-mils on 2-mil CuFlon <sup>®</sup>
Sensitivity to alignment within waveguide	High	Less than for E-plane probe

## V. Conclusions

Our wirebond test results indicate that excellent electrical performance is possible using CPWG wirebond interfaces above 100 GHz provided that extra care is taken to minimize bond length and faithfully reproduce the CPW trace geometry with the bonds. The motherboard / daughterboard packaging approach exploits this minimum length low loss wirebond interface to facilitate evaluation and characterization of a complex chain of MMICs. The design presented here is perhaps most applicable for development and prototyping, as well as preparation of MMIC sub-assemblies for system insertion. Finally, the performance of E-plane probe and antipodal finline transitions for use in G-band waveguide split block packages have been shown to exhibit reasonable electrical characteristics. It has been shown that the van Heuven transition fabricated on Cuflon<sup>®</sup> is a viable approach for incorporation in a split-block waveguide package applicable for G-band operation. We anticipate that further scaling of all of these approaches to higher frequencies is possible (at least through the 220-325 GHz Y-band), though many predominantly mechanical difficulties related to thinner substrates, finer trace geometries, and tighter alignment tolerances may ultimately force less conventional approaches to be used, such as the membrane methods used with THz frequency mixers [1].

## References

- [1] P. H. Siegel, R. P. Smith, M. Gaidis, S. Martin, J. Podosek, U. Zimmermann, "2.5 THz monolithic mebrane-diode mixer," Proceedings of Ninth Intl. Symp. Space THz Technology, pp. 147-159, March 1998.
- [2] C. M. Mann, D. N. Matheson, et. al., "On the design and measurement of a 2.5 THz waveguide mixer," Proceedings of Ninth Intl. Symp. Space THz Technology, pp. 161-171, March 1998.

- [3] J. H. den Besten, D. Caprioli, et. al., "Coplanar waveguides and butt-joints on InP," Proceedings of 6<sup>th</sup> Annual Symp. of IEEE/LEOS Benelux Chapter, December 2001.
- [4] T. P. Budka, "Wide-bandwidth millimeter-wave bond-wire interconnects," IEEE Trans. Microwave Theory Tech., vol. 49, pp. 715-718, April 2001.
- [5] V. Radisic, L. Samoska, et. al., "80 GHz MMIC HEMT VCO," IEEE Microwave and Wireless Components Letters, Vol.11, pp. 325- 327, 2001.
- [6] V. Radisic, M. Micovic, et. al., "164-GHz MMIC HEMT doubler," IEEE Microwave and Wireless Components Letters, Vol.11, pp. 241-243, 2001.
- [7] J. Sor, Q. Yongxi, T. Itoh, "Coplanar waveguide fed quasi-Yagi antenna," Electronics Letters, Vol. 36, pp. 1-2, 2000.
- [8] L. Samoska, J. Bruston, A. Peralta, "Advanced HEMT MMIC circuits for millimeter- and submillimeter-wave power sources," Proceedings of the Far Infrared, Millimeter and Submillimeter-wave Detector Workshop, Monterey, CA, April 2002.
- [9] L. Samoska, et. al., "InP MMIC chip set covering 80-170 GHz," Proceeding of Twelfth Intl. Symp. Space THz Technology, Dec. 2001.
- [10] B. Glance, and R. Trambarulo, "A Waveguide to suspended stripline transition," IEEE Trans. Microwave Theory Tech., Vol. MTT-21, pp 117-118, Feb 1973.
- [11] Y.C. Shih, et al., "Waveguide-to-microstrip transitions for millimeter-wave applications," IEEE. Intl. Microwave Symp. Digest, pp. 473-475, June 1988.
- [12] Y.C. Leong, and S. Weinreb, "Full band waveguide-to-microstrip probe transitions," IEEE. Intl. Microwave Symp. Digest, pp. 1435-1438, June 1999.
- [13] J.H.C. van Heuven, "A new integrated waveguide-microstrip transition," IEEE Trans. Microwave Theory Tech., vol. MTT-24, pp. 144-147, March 1976.
- [14] V. Sokolov, P. Saunier, R. C. Bennett, and R. E. Lehmann, "20 GHz multistage FET power amplifiers," 1981 Telemetry Conference Proceedings, pp. 845-853, October 1981.

### **Acknowledgements**

This effort was supported in part by DARPA/MTO, through Contract N66001-99-CC-8605 from SPAWAR Systems Center San Diego. The authors would like to thank T. A. Funk, D. K. Jensen, E. M. Doherty, and S. J. Richardson for manuscript and artwork preparation.